

Interference suppression with iterative channel and spatial covariance matrix estimation for LTE downlink



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ABSTRACT

In LTE downlink systems with a high frequency reuse factor, inter-cell interference has been viewed as a predominant factor limiting system performance. Therefore, we develop an iterative interference suppression receiver to mitigate inter-cell interference for LTE downlink systems in this paper. The key challenge here is how to accurately estimate the channel state information and spatial interference covariance matrix by using limited available pilots to improve the interference suppression capability and the successful probability of signal detection. To overcome the challenge, an iterative receiver making full use of the time–frequency correlation of channel is proposed. Simulation results show that the proposed iterative receiver can obtain a lower packet-error ratio compared with the existing algorithms.

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1. Introduction

In conventional cellular systems, the users at different locations from the same cell or from the different cells may employ the same frequency, which will inevitably result in intra-cell or inter-cell interference, respectively. The former can be generally avoided or mitigated by carefully designing transmission signals or allocating resource while the latter is hard to deal with if there is no cooperation among multiple cells [1]. Consequently, the inter-cell interference has been viewed as a predominant limiting factor for the performance of cellular systems, especially for the cellular systems with a high frequency reuse factor, such as *long term evolution* (LTE) systems with full frequency reuse [1]. In this case, some effective schemes have been proposed to mitigate interference, which mainly include resource allocation and preprocessing at the transmitter side to avoid interference [1] and the interference suppression at the receiver side [2–11] to mitigate interference. In this paper, we mainly focus on the interference suppression at the receiver side for LTE downlink.

For the interference suppression receiver design, a basic idea is first to estimate the channel state information and interference covariance matrix based on available pilots, and then to implement interference suppression and signal detection by using the estimated channel and the spatial covariance matrix [2–11]. This means that the interference suppression capability of receiver will strictly depend on the accuracy of channel state information and spatial interference covariance estimations. If the channel state information is known, the receiver design only need to focusing on the signal detection, such as literatures [2–7]. Based on the known

channel information, the authors in [2] first use turbo principle to design a soft-input soft-output *minimum mean-square error* (MMSE) detector for code division multiple access (CDMA) systems while the authors in [3] use the iterative successive parallel arbitrated *decision feedback* (DF) to develop an MMSE-DF receiver for direct sequence CDMA systems. Furthermore, the authors in [4–6] apply the turbo principle to MIMO systems and developed several efficient iterative MIMO receivers. Considering the time variation of the practical MIMO channel in a packet, a joint tracking and equalization algorithm based on a Kalman filter is proposed in [7]. Depending on the iterative process, the performance of the aforementioned receivers is continuously improved.

However, the channel state information in practical systems is unknown and need to be estimated. Generally, it is difficult to obtain an accurate estimation of channel and spatial covariance matrix because the available pilots are limited and the interference is changed with the scheduled period in practical systems, such as in LTE systems [1]. To overcome this difficulty, several channel and spatial covariance estimation schemes which can be applied to LTE systems have been proposed in [8–11]. In [8], a time-domain low-pass weighting filter has been proposed to improve spatial correlation matrix estimation of interference, which is mainly based on the Cholesky decomposition to exploit channel frequency-domain correlation to smooth the noise. Unfortunately, this leads to the channel spectral leakage in practical systems so that the estimation performance is degraded. Therefore, a time-domain sinc type weighing scheme with the frequency-domain moving average filter is used in [9–11] to reduce the estimation error caused by the spectral leakage. The channel in practical LTE systems is generally constituted by the time- and frequency-domain correlations, but the aforementioned schemes only use one part of both channel correlations. Therefore, we will propose an iterative interference

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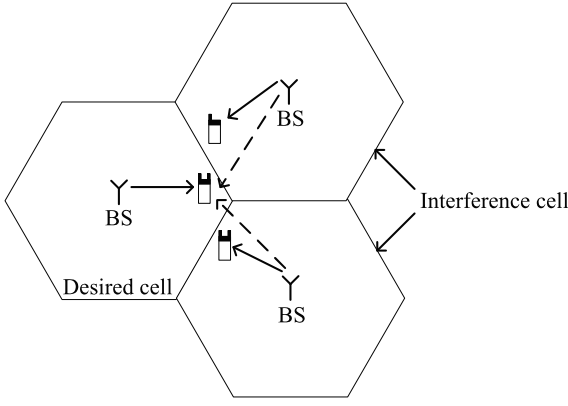


Fig. 1. Multi-cell interference model.

suppression receiver with enhanced channel estimation and improved spatial covariance matrix estimation in this paper. Its key idea is to make full use of the time–frequency correlations of channel to improve the estimation accuracy of channel state information and spatial covariance matrix in each iteration. Meanwhile, it can also use the iterative processing to continuously improve the performance of the proposed receiver.

The rest of this paper is organized as follows. Section 2 introduces the signal model and the channel model in LTE downlink systems. In Section 3, an iterative receiver with enhanced channel estimation and spatial covariance matrix estimation is proposed. Simulation results are presented in Section 4 and conclusions are drawn in Section 5.

Here are some notations used in this paper.

- $(\cdot)^H$: the Hermitian operation;
- $\mathbb{E}\{\cdot\}$: the expectation operation;
- $\text{trace}\{\cdot\}$: the trace operator;
- $\text{diag}\{\cdot\}$: the diagonal operation;
- $\mathbf{I}_m$: the $m \times m$ identity matrix;
- $\mathbf{F}_m$: the $m \times m$ Fourier transform matrix;
- $[\mathbf{A}]_{n,m}$: the element at the n th row and the m th column of matrix $\mathbf{A}$;
- $\det(\mathbf{A})$: the determinant of matrix $\mathbf{A}$;
- $1:l$: from 1 to l .

2. System model

In this section, we first describe the signal model and then discuss channel model and pilot structure in LTE downlink.

2.1. Signal model and problem description

As shown in Fig. 1, we consider a downlink multi-cell system consisting of N_{cells} cells, which share the same available spectrum, i.e., the frequency reuse factor is one. Each cell has a base station (BS) with N_r receive antennas and U users each with N_t transmit antennas. At the BS, the scheduler will select users and allocate them to a subset of resource blocks (RB's). For orthogonal frequency division multiple access (OFDMA) exploited in LTE downlink, each user within the same cell will be orthogonally allocated a different subset of RB's. Therefore, the intra-cell interference is avoided, but the inter-cell interference still exists if there is no coordination among different cells. In this case, the received signal at the k th subcarrier of the n th OFDM symbol can be expressed as

$$\mathbf{y}_{n,k} = \mathbf{H}_{n,k} \mathbf{x}_{n,k} + \mathbf{z}_{n,k}, \quad (1)$$

where $\mathbf{H}_{n,k}$ is an $N_r \times N_t$ channel matrix at the k th subcarrier of the n th OFDM symbol, $\mathbf{x}_{n,k}$ is an $N_t \times 1$ transmit signal vector at

the k th subcarrier of the n th OFDM symbol, $\mathbf{z}_{n,k}$ is an $N_r \times 1$ noise and interference vector. The noise and interference vector, $\mathbf{z}_{n,k}$, is assumed with zero mean and covariance matrix, $\mathbf{R}_{n,k}$, at the k th subcarrier of the n th OFDM symbol. The spatial covariance matrix of interference and noise, $\mathbf{R}_{n,k}$ can be expressed as

$$\mathbf{R}_{n,k} = \mathbb{E}\{\mathbf{z}_{n,k} \mathbf{z}_{n,k}^H\} = \begin{bmatrix} r_{n,k}^{(11)} & \cdots & r_{n,k}^{(1N_r)} \\ \vdots & \ddots & \vdots \\ r_{n,k}^{(N_r 1)} & \cdots & r_{n,k}^{(N_r N_r)} \end{bmatrix}, \quad (2)$$

where $r_{n,k}^{(n_2 n_1)}$ denotes the correlation coefficient of the noise and interference from different received antennas, n_1 and n_2 . It should be noted that we assume the transmit signal, the interference, and the noise are independent each other in this paper.

Based on the Gaussian approximation [12], the probability density function (PDF) of the received signal in (1) can be expressed as

$$\begin{aligned} p(\mathbf{y}_{n,k} | \mathbf{x}_{n,k}, \mathbf{H}_{n,k}, \mathbf{R}_{n,k}) &= \frac{1}{\det(\pi \mathbf{R}_{n,k})} e^{-(\mathbf{y}_{n,k} - \mathbf{H}_{n,k} \mathbf{x}_{n,k})^H \mathbf{R}_{n,k}^{-1} (\mathbf{y}_{n,k} - \mathbf{H}_{n,k} \mathbf{x}_{n,k})} \\ &= \frac{1}{\det(\pi \mathbf{R}_{n,k})} e^{-\|(\mathbf{R}_{n,k}^{-1/2})^H (\mathbf{y}_{n,k} - \mathbf{H}_{n,k} \mathbf{x}_{n,k})\|^2}. \end{aligned} \quad (3)$$

From the above equation, the maximum likelihood detection (MLD) can be used to estimate the transmitted signal, $\mathbf{x}_{n,k}$, if the spatial covariance matrix, $\mathbf{R}_{n,k}$, and the channel matrix, $\mathbf{H}_{n,k}$, are known. However, $\mathbf{R}_{n,k}$ and $\mathbf{H}_{n,k}$ in (3) are generally unknown and need to be estimated by using pilot signals or training symbols in practical systems. Therefore, joint estimation for $\mathbf{x}_{n,k}$, $\mathbf{R}_{n,k}$, and $\mathbf{H}_{n,k}$ is required to maximize the PDF, $p(\mathbf{y}_{n,k} | \mathbf{x}_{n,k}, \mathbf{H}_{n,k}, \mathbf{R}_{n,k})$. Since this joint estimation has very high complexity, we will develop an iterative scheme with low complexity in Section 3 to simplify it.

2.2. Channel model and pilot structure

To obtain a good result of the channel estimation and the spatial covariance matrix estimation, we will first discuss the channel model and the available pilot structure for LTE downlink in this section.

The impulse response of equivalent baseband channel model from the n_t th transmit antenna to the n_r th receive antenna can be described by

$$h_{t,\tau}^{(n_r n_t)} = \sum_l \alpha_{t,l}^{(n_r n_t)} \delta(\tau - \tau_l^{(n_r n_t)}), \quad (4)$$

where $\tau_l^{(n_r n_t)}$ and $\alpha_{t,l}^{(n_r n_t)}$ are the delay and complex gain of the l th distinguishable path, respectively. Correspondingly, the frequency-domain response of the baseband channel in (4) can be expressed as

$$H_{t,f}^{(n_r n_t)} = \sum_l \alpha_{t,l}^{(n_r n_t)} \exp(-j2\pi f \tau_l^{(n_r n_t)}). \quad (5)$$

Since each distinguishable path consists of many unresolvable sub-paths, it is usually modeled as a wide-sense stationary (WSS) narrowband complex Gaussian processes with the same normalized correlation function. Different transmit–receive antenna pairs are assumed to have the same power–delay profile due to the small antenna space in centralized multi-antenna systems [14]. As a result, the time–frequency correlation of the channel can be expressed as [13]

$$r_H(\Delta f, \Delta t) = \sigma_H^2 r_t(\Delta t) r_f(\Delta f), \quad (6)$$

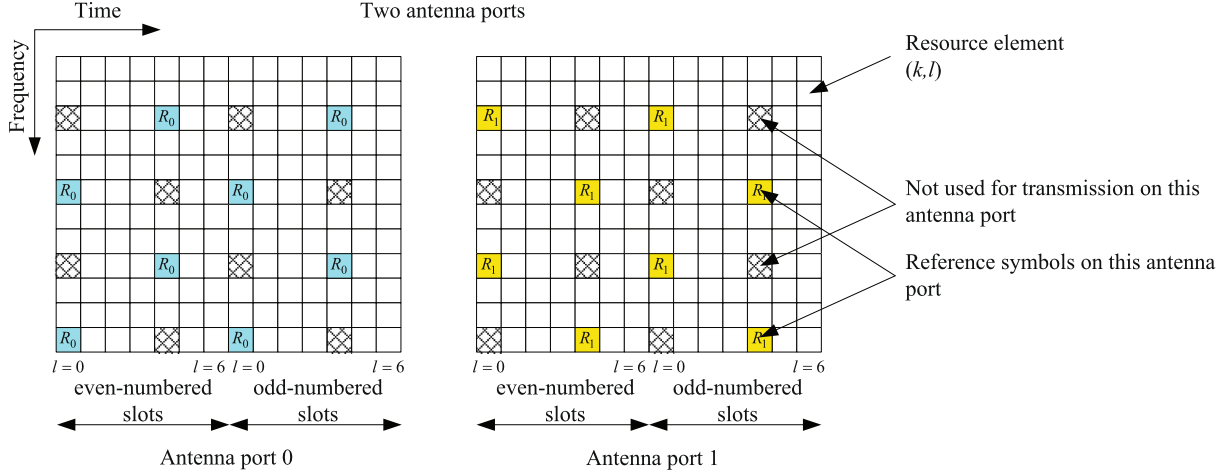


Fig. 2. Cell-specific pilot structure of two antenna port with normal cyclic prefix [15].

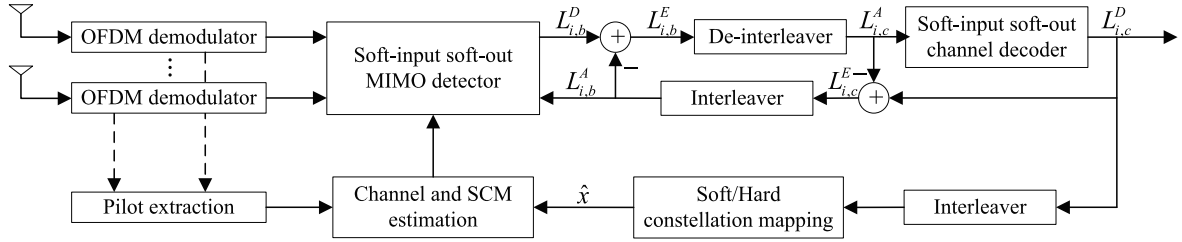


Fig. 3. Iterative receiver with channel and spatial covariance estimation.

where $r_t(\Delta t)$ is the time-domain correlation function determined by the Doppler spectrum, $r_f(\Delta f)$ is the frequency-domain correlation function determined by power delay profile and is defined as $r_f(\Delta f) = \sum_l \frac{\sigma_l^2}{\sigma_H^2} \exp(-j2\pi f \tau_l)$, σ_l^2 is the power of the l th distinguishable path, $\sigma_H^2 = \sum_{l=0}^{L-1} \sigma_l^2$ is the total average power of the time-domain channel, L is the length of the time-domain channel, i.e., the number of the distinguishable channel paths.

Assume that there are K subcarriers in an OFDM symbol and the subcarrier space is Δf , then the duration of each OFDM symbol and the sampling interval are $T_s = 1/\Delta f$ and $\frac{1}{K\Delta f}$, respectively. Consequently, the equivalent discrete complex baseband channel can be expressed as

$$H_{n,k}^{(n_r n_t)} = \frac{1}{\sqrt{K}} \sum_{l=0}^{L-1} \alpha_{n,l}^{(n_r n_t)} \exp\left(-j \frac{2\pi k \tau_l}{T_s}\right), \quad (7)$$

where $H_{n,k}^{(n_r n_t)}$ is an element from the n_r -th row and the n_t -th column of $\mathbf{H}_{n,k}$ in (1).

The distinguishable paths in practical systems are generally sparse [19,20], i.e., the time-domain channel has only limited non-zero paths. Therefore, we can exploit this fact to improve the estimation accuracy of the channel and spatial covariance matrices.

Depending on the channel statistics in LTE systems, four different pilot structures are defined for the downlink [15]. In this paper, we only exploit the cell-specific pilot structure, as shown in Fig. 2, to estimate the channel matrix and the spatial covariance matrix. For the cell-specific pilots, each RB only includes four pilots corresponding to each antenna port. Therefore, a key problem is how to improve the channel and spatial covariance matrix estimation by using the limited pilots in each scheduled duration since different scheduled durations maybe face different inter-cell interference.

3. Iterative receiver with interference suppression

To maximize the PDF, $p(\mathbf{y}_{n,k} | \mathbf{x}_{n,k}, \mathbf{H}_{n,k}, \mathbf{R}_{n,k})$, in (3), we need to jointly estimate the channel matrix, the spatial covariance matrix, and the transmit signal. However, the complexity of joint estimation is too high, we therefore develop a low-complexity iterative estimation scheme as shown in Fig. 3, which mainly includes three parts: the channel estimation, the spatial covariance matrix estimation, and the signal detection. In this scheme, we mainly use the time-frequency correlation of the channel to improve the accuracy of the channel and spatial covariance matrix estimation, and then the enhanced channel estimation and the improved spatial covariance matrix estimation is used to improve the interference suppression capability and the successful probability of signal detection.

3.1. Channel estimation

In this section, we will develop a practical enhanced channel estimation scheme, which mainly uses the time-frequency correlation to improve the estimation accuracy.

To the best of our knowledge, if the statistical information of the channel, the interference, and the noise is known, then MMSE estimator can be used to achieve the optimal estimation performance [13,14]. However, the other statistical information except the noise is hardly obtained in practical systems. Therefore, *least square* (LS) estimator is generally used to obtain an initial estimation result, and then this result can be used to estimate the corresponding statistical parameters to further improve channel estimation accuracy.

Assume that $\mathcal{K}_p$, $\mathcal{K}_d$, and $\mathcal{K}_u$ are the index sets of the frequency-domain pilot subcarriers, data subcarriers, and unavailable high frequency and direct current subcarriers for LTE downlink, respectively. Denote $\mathcal{N}_p$ and $\mathcal{N}_d$ to be the index sets of the OFDM symbols with and without pilots for LTE downlink, respec-

tively. Then, LS estimation of the channel vector corresponding to the n_t th transmit antenna can be expressed as [13,14]

$$\hat{\mathbf{H}}_{n,k}^{(n_t)} = \frac{\mathbf{y}_{n,k}}{\mathbf{x}_{n,k}^{(n_t)}} = \mathbf{H}_{n,k}^{(n_t)} + \frac{\mathbf{z}_{n,k}}{\mathbf{x}_{n,k}^{(n_t)}} = \mathbf{H}_{n,k}^{(n_t)} + \hat{\mathbf{z}}_{n,k}, \quad (8)$$

where $k \in \mathcal{K}_p$ and $n \in \mathcal{N}_p$ for the initial channel estimation stage while $k \in \mathcal{K}_p \cup \mathcal{K}_d$ and $n \in \mathcal{N}_p \cup \mathcal{N}_d$ for the iterative channel estimation stage, $\hat{\mathbf{H}}_{n,k}^{(n_t)}$ denotes the channel vector from the n_t th transmit antenna to all receive antennas in the desired cell and can be viewed as the n_t th column of $\hat{\mathbf{H}}_{n,k}$. In the following, the initial channel estimation and the iterative channel estimation will be identically performed. For simplifying expressions, we denote the index set of all subcarriers corresponding to known available signals and the index set of all OFDM symbols corresponding to known available signals are defined as

$$\mathcal{K}_a \triangleq \begin{cases} \mathcal{K}_p, & \text{for initial estimation,} \\ \mathcal{K}_p \cup \mathcal{K}_d, & \text{for iterative estimation,} \end{cases}$$

and

$$\mathcal{N}_a \triangleq \begin{cases} \mathcal{N}_p, & \text{for initial estimation,} \\ \mathcal{N}_p \cup \mathcal{N}_d, & \text{for iterative estimation,} \end{cases}$$

respectively, where the available known signals include pilots or/and estimation signals.

After the channel state information at the known subcarriers, which are either the pilot subcarriers for the initial estimation or the pilot and data subcarrier for the iterative estimation, is obtained, the channel state information at the other subcarriers can be calculated by using the interpolation. The two-dimension MMSE channel interpolation for grid pilots has a very good performance [16], but it requires channel statistics and is very complex. Therefore, we will propose a robust and low-complexity channel interpolation scheme consisting of the cascade of two one-dimension interpolations.

Assume $\mathbf{B}$ is a frequency-domain interpolation matrix, then the channel vector on all subcarriers of the n th OFDM symbol from the n_t th transmit antenna to the n_r th receive antenna, $\mathbf{H}_{n,\mathcal{K}}^{(n_r n_t)} = [H_{n,1}^{(n_r n_t)} \ H_{n,2}^{(n_r n_t)} \ \dots \ H_{n,K}^{(n_r n_t)}]^T$, can be estimated by

$$\tilde{\mathbf{H}}_{n,\mathcal{K}}^{(n_r n_t)} = \mathbf{B}\hat{\mathbf{H}}_{n,\mathcal{K}_a}^{(n_r n_t)} = \mathbf{H}_{n,\mathcal{K}}^{(n_r n_t)} + \tilde{\mathbf{z}}_{n,\mathcal{K}}, \quad (9)$$

where $\mathcal{K} = \mathcal{K}_p \cup \mathcal{K}_d \cup \mathcal{K}_u$ denotes the set of all subcarriers, while $\hat{\mathbf{H}}_{n,\mathcal{K}_a}^{(n_r n_t)}$ denotes a channel column vector from all known subcarriers, $\mathcal{K}_a$.

From (8) and (9), we can express the estimation mean-square error of the frequency-domain channel vector from the n_t th transmit antenna to the n_r th receive antenna at the n th OFDM symbol as

$$\begin{aligned} \varepsilon_f &= \mathbb{E}\{(\mathbf{H}_{n,\mathcal{K}}^{(n_r n_t)} - \mathbf{B}\hat{\mathbf{H}}_{n,\mathcal{K}_a}^{(n_r n_t)})^H (\mathbf{H}_{n,\mathcal{K}}^{(n_r n_t)} - \mathbf{B}\hat{\mathbf{H}}_{n,\mathcal{K}_a}^{(n_r n_t)})\} \\ &= \mathbb{E}\{\text{trace}\{(\mathbf{H}_{n,\mathcal{K}}^{(n_r n_t)} - \mathbf{B}\hat{\mathbf{H}}_{n,\mathcal{K}_a}^{(n_r n_t)})(\mathbf{H}_{n,\mathcal{K}}^{(n_r n_t)} - \mathbf{B}\hat{\mathbf{H}}_{n,\mathcal{K}_a}^{(n_r n_t)})^H\}\} \\ &= \mathbb{E}\{\text{trace}\{((\mathbf{F}_{\mathcal{K},\mathcal{L}} - \mathbf{B}\mathbf{F}_{\mathcal{K}_a,\mathcal{L}})\mathbf{h}_n^{(n_r n_t)} - \mathbf{B}\hat{\mathbf{z}}_{n,\mathcal{K}_a}) \\ &\quad \times ((\mathbf{F}_{\mathcal{K},\mathcal{L}} - \mathbf{B}\mathbf{F}_{\mathcal{K}_a,\mathcal{L}})\mathbf{h}_n^{(n_r n_t)} - \mathbf{B}\hat{\mathbf{z}}_{n,\mathcal{K}_a})^H)\}\} \\ &= \text{trace}\{(\mathbf{F}_{\mathcal{K},\mathcal{L}} - \mathbf{B}\mathbf{F}_{\mathcal{K}_a,\mathcal{L}})\mathbf{C}_h(\mathbf{F}_{\mathcal{K},\mathcal{L}} - \mathbf{B}\mathbf{F}_{\mathcal{K}_a,\mathcal{L}})^H \\ &\quad + \mathbf{B}\mathbf{T}_{n,\mathcal{K}_a}\mathbf{B}^H\}, \end{aligned} \quad (10)$$

where $\mathbf{C}_h = \mathbb{E}\{\mathbf{h}_n^{(n_r n_t)} \mathbf{h}_n^{(n_r n_t)H}\}$ contains the statistics of the time-domain channel from the n_t th transmit antenna to the n_r th receive antenna at the n th OFDM symbol and $\mathbf{T}_{n,\mathcal{K}_a} = \mathbb{E}\{\hat{\mathbf{z}}_{n,\mathcal{K}_a} \hat{\mathbf{z}}_{n,\mathcal{K}_a}^H\}$ is the frequency-domain statistic of interference and noise at all

the known subcarriers. As mentioned before, different transmit-receive antenna pairs have the same channel statistics, then, the interference channels to the different receiver antennas will have the same statistics. We therefore have the same $\mathbf{C}_h$ for all antenna pairs and the same $\mathbf{T}_{n,\mathcal{K}_p}$ at the different receive antennas.

By minimizing the mean-square error in (10), we have the optimal frequency-domain interpolation matrix as

$$\begin{aligned} \mathbf{B} &= \mathbf{F}_{\mathcal{K},\mathcal{L}} \mathbf{C}_h \mathbf{F}_{\mathcal{K}_a,\mathcal{L}}^H (\mathbf{F}_{\mathcal{K}_a,\mathcal{L}} \mathbf{C}_h \mathbf{F}_{\mathcal{K}_a,\mathcal{L}}^H + \mathbf{T}_{n,\mathcal{K}_a})^{-1} \\ &= \mathbf{F}_{\mathcal{K},\mathcal{L}} (\mathbf{C}_h^{-1} + \mathbf{F}_{\mathcal{K}_a,\mathcal{L}}^H \mathbf{T}_{n,\mathcal{K}_a}^{-1} \mathbf{F}_{\mathcal{K}_a,\mathcal{L}})^{-1} \mathbf{F}_{\mathcal{K}_a,\mathcal{L}}^H \mathbf{T}_{n,\mathcal{K}_a}^{-1}, \end{aligned} \quad (11)$$

where $\mathbf{F}_{\mathcal{K},\mathcal{L}}$ denotes a submatrix of Fourier matrix, $\mathbf{F}$, with subscripts $\mathcal{K}$ and $\mathcal{L}$, $\mathcal{L}$ denotes a set of non-zero channel paths. It should be noted that the second equality uses the matrix inverse theorem and we assume the correlation matrix of the time-domain channel, $\mathbf{C}_h$ and the correlation matrix of the interference and noise, $\mathbf{T}_{n,\mathcal{K}_a}$ are reversible. Theoretically, the interpolation matrix in (11) is optimal, but it is very hard to apply this scheme to practical LTE systems since obtaining the second-order statistical characteristic of the channel is difficult and a lot of high frequency subcarriers are unused in LTE systems. In fact, the frequency-domain correlation of the channel is mainly determined by the limited length of the time-domain channel [14]. Therefore, we will simplify the interpolation matrix in (11) as

$$\mathbf{B} = \mathbf{F}_{\mathcal{K},\hat{\mathcal{L}}} \mathbf{I}_{\hat{\mathcal{L}},\mathcal{K}} \mathbf{F}_{\mathcal{K},\mathcal{K}}^H \mathbf{U}, \quad (12)$$

where $\mathbf{U}$ is a linear interpolation matrix, $\mathbf{I}_{\hat{\mathcal{L}},\mathcal{K}}$ is a selection matrix with subscripts $\hat{\mathcal{L}}$ and $\mathcal{K}$, and $\hat{\mathcal{L}}$ is the index set of the selected non-zero channel coefficients. In fact, the interpolation matrix in (11) is mainly implemented by the following four steps: (1) the estimated channel at the known subcarriers is interpolated by using the linear filter, $\mathbf{U}$; (2) the frequency-domain channel based on linear interpolation is transformed into the time-domain one; (3) the non-zero time-domain channel coefficients are selected according to the set $\hat{\mathcal{L}}$; (4) the selected time-domain channel is transformed into the frequency-domain one. Obviously, the final interpolation result mainly depends on the selection of the non-zero time-domain channel coefficients, i.e., the determination of the set $\hat{\mathcal{L}}$. Since the estimation channel is a superposition of the perfect channel and the noise, the selected scheme can be determined by

$$\hat{\mathcal{L}} = \{\hat{l} \mid |\mathbf{F}_{\mathcal{K},\hat{l}}^H \mathbf{U} \hat{\mathbf{H}}_{n,\mathcal{K}_p}^{(n_r n_t)}|^2 > \rho \sigma_n^2, \ 1 \leq l \leq K\}, \quad (13)$$

where ρ is a proportional factor to select the non-zero channel paths and σ_n^2 is the noise covariance. Essentially, this frequency-domain channel estimation scheme mainly uses the limited length of the time-domain channel to improve the channel estimation accuracy.

In the above, we have obtained the frequency-domain channel estimation. In the following, we will develop a time-domain channel interpolation scheme. Assume that the time-domain channel interpolation matrix is $\mathbf{A}$, then the channel vector of all OFDM symbols at the k th subcarrier from the n_t th transmit antenna to the n_r th receive antenna, $\mathbf{H}_{\mathcal{N},k}^{(n_r n_t)} = [H_{1,k}^{(n_r n_t)} \ H_{2,k}^{(n_r n_t)} \ \dots \ H_{|\mathcal{N}|,k}^{(n_r n_t)}]^T$ can be estimated by

$$\tilde{\mathbf{H}}_{\mathcal{N},k}^{(n_r n_t)} = \mathbf{A} \tilde{\mathbf{H}}_{\mathcal{N}_a,k}^{(n_r n_t)}, \quad (14)$$

where $\mathcal{N}$ denotes a set of all OFDM symbols in a transmission time interval and $|\mathcal{N}|$ denotes its cardinality, $\tilde{\mathbf{H}}_{\mathcal{N}_a,k}^{(n_r n_t)}$ denotes a column channel vector from all known symbols. Similarly, by using the criterion of the mean square-error minimization, we can obtain the time-domain interpolation matrix as

$$\mathbf{A} = \mathbf{C}_H \mathbf{I}_{\mathcal{N}_a, \mathcal{N}}^H \mathbf{I}_{\mathcal{N}_a, \mathcal{N}} (\mathbf{C}_H + \mathbf{C}_z)^{-1} \mathbf{I}_{\mathcal{N}_a, \mathcal{N}}^H, \quad (15)$$

where $\mathbf{C}_H = \mathbb{E}\{\mathbf{H}_{\mathcal{N},k}^{(n_r n_t)} \mathbf{H}_{\mathcal{N},k}^{(n_r n_t)H}\}$ corresponds the channel time-varying characteristic, $\mathbf{C}_z = \mathbb{E}\{\mathbf{z}_{\mathcal{N},k}^{(n_r n_t)} \mathbf{z}_{\mathcal{N},k}^{(n_r n_t)H}\}$ reflects the statistic of frequency-domain channel estimation error, $\mathbf{I}_{\mathcal{N}_a, \mathcal{N}}$ is a selection matrix with $\mathbf{I}_{\mathcal{N}_a, \mathcal{N}} \mathbf{I}_{\mathcal{N}_a, \mathcal{N}}^H = \mathbf{I}$. Based on (15), the channel at the subcarrier k corresponding to the different OFDM symbols will be obtained.

As mentioned before, it is difficult to get second-order statistics of the channel, we will therefore propose a robust time-domain channel estimation scheme. Assume that the carrier frequency offset due to the oscillator mismatch between the transmitters and the receivers has been compensated, then the channel time-varying characteristics will be mainly determined by the Doppler spectrum. Consequently, we can approximately express $\mathbf{C}_H$ as [13]

$$\mathbf{C}_H = \begin{bmatrix} c_H[0] & c_H[1] & \cdots & c_H[|\mathcal{N}| - 1] \\ c_H[1] & c_H[0] & \cdots & c_H[|\mathcal{N}| - 2] \\ \vdots & \vdots & \ddots & \vdots \\ c_H[|\mathcal{N}| - 1] & c_H[|\mathcal{N}| - 2] & \cdots & c_H[0] \end{bmatrix}, \quad (16)$$

where $c_H[n] = \frac{\sin(n\omega_d)}{n\omega_d}$, $\omega_d = \frac{2\pi f_d}{\Delta f}$, f_d is the maximal Doppler offset determined by the mobile speed. Because the mobile speed of each user can be estimated easily, we assume that the user speed is known. Meanwhile, for simplicity, we can approximate $\mathbf{C}_z$ as

$$\mathbf{C}_z \approx \sigma_n^2 \mathbf{I}. \quad (17)$$

Obviously, the mobile speed of each user and the noise covariance can be easily estimated, therefore, this scheme can be easily applied to practical LTE systems.

3.2. Spatial covariance matrix estimation

In the last section, the channel estimation accuracy is enhanced by using the time–frequency correlation characteristics of the channel. Can we use this characteristics to improve the accuracy of the spatial covariance matrix estimation? To answer this question, we will first analyze the time–frequency correlation characteristics of the spatial covariance matrix. If the answer is yes, we can further use it to improve the accuracy of the spatial covariance matrix estimation.

3.2.1. Analysis of the spatial covariance matrix

Without loss of generality, we assume after CP of the received signal is removed, the frequency-domain expression of the interference from the other cells and noise at the n_t th receive antenna can be written as

$$\mathbf{z}_{n,K}^{(n_r)} = \sum_{n_l=1}^{N_l} \mathbf{F} \mathbf{G}_n^{(n_r n_l)} \mathbf{s}_n^{(n_l)} + \mathbf{v}_n^{(n_r)}, \quad (18)$$

where $\mathbf{v}_n^{(n_r)}$ is a noise vector with zero-mean and covariance matrix $\sigma_n^2 \mathbf{I}$,

$$\mathbf{s}_n^{(n_l)} = \begin{bmatrix} \underbrace{s_{n-1,m+1}^{(n_l)} \cdots s_{n-1,K}^{(n_l)}}_{K-m} \underbrace{s_{n,K-K_g+1}^{(n_l)} \cdots s_{n,K}^{(n_l)}}_{K_g} \underbrace{s_{n,1}^{(n_l)} \cdots s_{n,m}^{(n_l)}}_m \end{bmatrix}^T \quad (19)$$

is an asynchronous time-domain transmit signal from the n_l interference cell,

$$\mathbf{G}_n^{(n_r n_l)} = \begin{bmatrix} 0 & \cdots & 0 & g_{n,L-1}^{(n_r n_l)} & \cdots & g_{n,0}^{(n_r n_l)} & 0 & \cdots & \cdots & 0 \\ \vdots & \vdots & \vdots & 0 & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \vdots & \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \vdots & \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \vdots & \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & 0 & \cdots & \cdots & 0 & g_{n,L-1}^{(n_r n_l)} & \cdots & g_{n,0}^{(n_r n_l)} \end{bmatrix} \quad (20)$$

is the channel matrix from the BS of the n_l th interference cell to the n_r th receive antenna of the desired user, K , K_g , m , and N_l are the number of the subcarriers in an OFDM symbols, the CP length, the asynchronous time-offset, and the number of all interference cells, respectively. Obviously, if $m = K$, then the interference signal will synchronize with the desired signal.

From (18), we can express the channel correlation of arbitrary two receive antennas as

$$\begin{aligned} \mathbf{r}_{n,K}^{(n_r1 n_r2)} &= \text{diag}\{\mathbb{E}\{\mathbf{z}_{n,K}^{(n_r1)} \mathbf{z}_{n,K}^{(n_r2)H}\}\} \\ &= \text{diag}\left\{\mathbb{E}\left\{\left(\sum_{N_l=1}^{N_l} \mathbf{F} \mathbf{G}_n^{(n_r1 N_l)} \mathbf{s}_n^{(n_r1 N_l)} + \mathbf{v}_n^{(n_r1)}\right) \right. \right. \\ &\quad \times \left. \left(\sum_{N_l=1}^{N_l} \mathbf{F} \mathbf{G}_n^{(n_r2 N_l)} \mathbf{s}_n^{(n_r2 N_l)} + \mathbf{v}_n^{(n_r2)}\right)^H\right\}\} \\ &= \text{diag}\left\{\mathbf{F} \left(\sum_{n_l1=1}^{N_l} \sum_{n_l2=1}^{N_l} \mathbf{G}_n^{(n_r1 n_l1)} \mathbb{E}\{\mathbf{s}_n^{(n_r1 n_l1)} \mathbf{s}_n^{(n_r2 n_l2)H}\} \right. \right. \\ &\quad \times \left. \left. \mathbf{G}_n^{(n_r2 n_l2)}\right)^H \mathbf{F}^H + \sigma_n^2 \mathbf{I}\right\}. \end{aligned} \quad (21)$$

In practical LTE systems, the transmit signals from the different cells are independent each other, we therefore have

$$\mathbb{E}\{\mathbf{s}_n^{(n_r1 n_l1)} \mathbf{s}_n^{(n_r2 n_l2)H}\} = \begin{cases} \mathbf{I}, & n_l1 = n_l2, \\ \mathbf{0}, & n_l1 \neq n_l2. \end{cases} \quad (22)$$

Then from (21) and (22), we can simplify the channel correlation of arbitrary two antennas as

$$\mathbf{r}_{n,K}^{(n_r1 n_r2)} = \text{diag}\left\{\mathbf{F} \left(\sum_{n_l=1}^{N_l} \mathbf{G}_n^{(n_r1 n_l)} \mathbf{G}_n^{(n_r2 n_l)H}\right) \mathbf{F}^H + \sigma_n^2 \mathbf{I}\right\}. \quad (23)$$

From (23), we find that the element, $\mathbf{r}_{n,K}^{(n_r1 n_r2)}$, of the spatial covariance matrix is similar as the channel correlation. To show this fact more clearly, we give an example with $K = 512$, $K_g = 36$, $m = 256$, $\text{INR} = 10$ dB in Fig. 4, where INR denotes the interference-to-noise ratio. From this figure, we find that time-domain length of spatial interference covariance matrix is limited which is similar as the characteristic of time-domain channel no matter whether the interference are synchronous with the desired signals or not. It should be noted that in this figure, the frequency-domain zero-zone is because the high frequency subcarriers are unused while the time-domain zero-zone is because the time-domain channel has limited length. These analysis shows that we can also use the time–frequency correlation to further improve the accuracy of spatial covariance matrix estimation.

3.2.2. Improved spatial covariance matrix estimation

Once the channel information and the transmitted signal are given, the instantaneous spatial covariance matrix at the k th subcarrier of the n th OFDM symbol can be estimated by

$$\hat{\mathbf{R}}_{n,k} = (\mathbf{y}_{n,k} - \check{\mathbf{H}}_{n,k} \hat{\mathbf{x}}_{n,k})(\mathbf{y}_{n,k} - \check{\mathbf{H}}_{n,k} \hat{\mathbf{x}}_{n,k})^H. \quad (24)$$

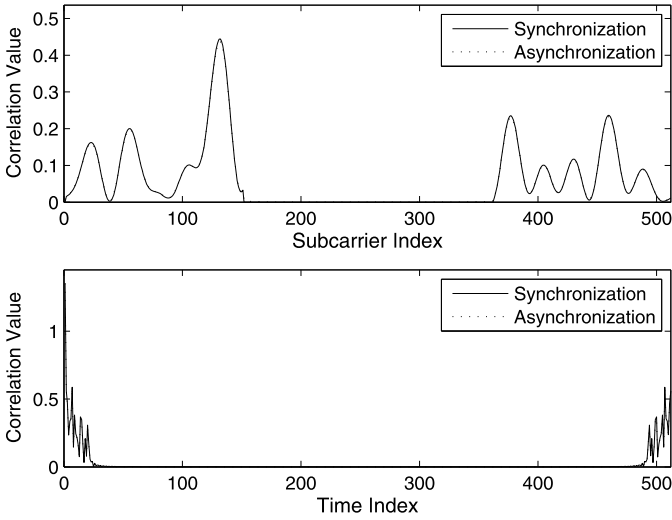


Fig. 4. The time–frequency correlation of spatial interference covariance matrix for LTE downlink with $K = 512$, $K_g = 36$, $m = 256$, $INR = 10$ dB.

Based on the instantaneous estimation in (24), several schemes in [8–11] have been proposed to obtain the spatial covariance matrix estimation.

In [8], the authors assume that all OFDM symbols have the same instantaneous spatial covariance matrix so that the statistical spatial covariance matrix can be estimated by

$$\tilde{\mathbf{R}}_k = \frac{1}{N} \sum_n \hat{\mathbf{R}}_{n,k}. \quad (25)$$

Based on the estimation results in (25), the Cholesky decomposition can be used to further improve the accuracy of the spatial covariance matrix of interference [8]. If the instantaneous spatial covariance matrices are different at different OFDM symbols but are very close at several adjacent subcarriers [9–11], the statistical spatial covariance matrix can be estimated by

$$\tilde{\mathbf{R}}_{n,k} = \sum_{q=-Q/2}^{Q/2} \alpha_q \hat{\mathbf{R}}_{n,k+q}, \quad (26)$$

where $\alpha_q = \frac{1}{Q}$ [9–11]. Then, the estimation results in (26) can be further smoothed to improve the estimation accuracy by using the frequency-domain correlation of channel. However, the aforementioned schemes only use one part of the time-domain correlation of channel and the frequency-domain correlation of channel. Therefore, we will simultaneously use the time–frequency correlation of channel in the iterative processing to improve the performance of receiver.

As mentioned before, the correlation characteristic of each transmit–receive antenna pair has a limited length. Then, we can use the similar idea as the channel estimation to improve the spatial covariance matrix estimation. Correspondingly, we can express the frequency-domain estimation as

$$\tilde{\mathbf{r}}_{n,\mathcal{K}}^{(n_{r1}n_{r2})} = \mathbf{B}\hat{\mathbf{r}}_{n,\mathcal{K}_a}^{(n_{r1}n_{r2})}, \quad (27)$$

where $\hat{\mathbf{r}}_{n,\mathcal{K}_a}^{(n_{r1}n_{r2})}$ is the correlation characteristic of the n_{r1} th receive antenna and the n_{r2} th receive antenna at the known subcarriers $\mathcal{K}_a$. Meanwhile, the time-domain estimation can be written by

$$\tilde{\mathbf{r}}_{\mathcal{N},k}^{(n_{r1}n_{r2})} = \mathbf{A}\hat{\mathbf{r}}_{\mathcal{N}_a,k}^{(n_{r1}n_{r2})}, \quad (28)$$

for $n_{r1} \geq n_{r2}$. While for $n_{r1} < n_{r2}$, we have

$$\tilde{\mathbf{r}}_{n,\mathcal{K}}^{(n_{r1}n_{r2})} = \tilde{\mathbf{r}}_{n,\mathcal{K}}^{(n_{r2}n_{r1})H}, \quad (29)$$

where $\tilde{\mathbf{r}}_{n,\mathcal{K}}^{(n_{r2}n_{r1})}$ can be obtained according to (28). Then from (28) and (29), the corresponding spatial covariance matrix can be estimated by

$$\tilde{\mathbf{R}}_{n,k} = \begin{bmatrix} \tilde{\mathbf{r}}_{n,k}^{(11)} & \cdots & \tilde{\mathbf{r}}_{n,k}^{(1N_r)} \\ \vdots & \ddots & \vdots \\ \tilde{\mathbf{r}}_{n,k}^{(1N_r)*} & \cdots & \tilde{\mathbf{r}}_{n,k}^{(N_rN_r)} \end{bmatrix} + \sigma_n^2 \mathbf{I}. \quad (30)$$

It should be noted that the filter matrices, $\mathbf{A}$ and $\mathbf{B}$, here have the same design methods as the channel estimation, but are with different filter parameters.

3.3. Signal detection

Once the estimation results of channel state information and spatial covariance matrix are obtained, we can estimate the transmitted signal by maximizing the condition PDF in (3). Consequently, the estimated signal can be expressed as

$$\hat{\mathbf{x}}_{n,k} = \min_{\mathbf{x}_{n,k} \in \mathcal{X}} \|(\mathbf{R}_{n,k}^{-1/2})^H (\mathbf{y}_{n,k} - \mathbf{H}_{n,k} \mathbf{x}_{n,k})\|^2, \quad (31)$$

where $\mathcal{X}$ denotes a set consisting of all possible transmitted signals, $\mathbf{x}_{n,k}$. In fact, the operation of $(\mathbf{R}_{n,k}^{-1/2})^H$ in (31) is to whiten the interference while improving the successful probability of signal detection. Therefore, we need to continuously improve the accuracy of the spatial covariance matrix in the iterative processes. Correspondingly, we can also express the bit *log-likelihood ratio* (LLR) of the estimated signal as

$$\hat{L}_{n,k}^{(m)} = \ln \frac{\sum_{\mathbf{x}_{n,k} \in \mathcal{X}_{m,+1}} e^{\lambda(\mathbf{x}_{n,k}|\mathbf{y}_{n,k}, \mathbf{R}_{n,k}^{-1/2}, \mathbf{H}_{n,k})}}{\sum_{\mathbf{x}_{n,k} \in \mathcal{X}_{m,-1}} e^{\lambda(\mathbf{x}_{n,k}|\mathbf{y}_{n,k}, \mathbf{R}_{n,k}^{-1/2}, \mathbf{H}_{n,k})}}, \quad (32)$$

where $\mathcal{X}_{m,\pm 1}$ denote the sets of all possible bit vectors with the m th element as ± 1 and

$$\lambda(\mathbf{x}_{n,k}|\mathbf{y}_{n,k}, \mathbf{R}_{n,k}^{-1/2}, \mathbf{H}_{n,k}) = \|(\mathbf{R}_{n,k}^{-1/2})^H (\mathbf{y}_{n,k} - \mathbf{H}_{n,k} \mathbf{x}_{n,k})\|^2.$$

When performing iteration, we can use the initial LLR estimation, $\hat{L}_{n,k}^{(m)}$, as an a priori information to update a posteriori LLR information, $L_{n,k}^{D_1(m)}$, which can be expressed as [17,18]

$$\hat{L}_{n,k}^{D_1(m)} = \hat{L}_{n,k}^{A_1(m)} + \ln \frac{\sum_{\mathbf{x}_{n,k} \in \mathcal{X}_{m,+1}} e^{\lambda(\mathbf{x}_{n,k}, \mathbf{x}_{n,k,[m]}, \hat{\mathbf{L}}_{n,k,[m]}^{D_1} | \mathbf{y}_{n,k}, \mathbf{R}_{n,k}^{-1/2}, \mathbf{H}_{n,k})}}{\sum_{\mathbf{x}_{n,k} \in \mathcal{X}_{m,-1}} e^{\lambda(\mathbf{x}_{n,k}, \mathbf{x}_{n,k,[m]}, \hat{\mathbf{L}}_{n,k,[m]}^{D_1} | \mathbf{y}_{n,k}, \mathbf{R}_{n,k}^{-1/2}, \mathbf{H}_{n,k})}}, \quad (33)$$

$\underbrace{\hspace{10em}}_{\hat{L}_{n,k}^{E_1(m)}}$

where

$$\begin{aligned} \lambda(\mathbf{x}_{n,k}, \mathbf{x}_{n,k,[m]}, \hat{\mathbf{L}}_{n,k,[m]}^{D_1} | \mathbf{y}_{n,k}, \mathbf{R}_{n,k}^{-1/2}, \mathbf{H}_{n,k}) \\ = -\|(\mathbf{R}_{n,k}^{-1/2})^H (\mathbf{y}_{n,k} - \mathbf{H}_{n,k} \mathbf{x}_{n,k})\|^2 + \frac{1}{2} \mathbf{x}_{n,k,[m]}^T \hat{\mathbf{L}}_{n,k,[m]}^{D_1} \mathbf{x}_{n,k,[m]}, \end{aligned}$$

$\hat{L}_{n,k}^{E_1(m)}$ is the extrinsic information, $\mathbf{x}_{n,k,[m]}$ is the subvector of $\mathbf{x}_{n,k}$ which is obtained by removing the m th element of $\mathbf{x}_{n,k}$, and $\hat{\mathbf{L}}_{n,k,[m]}^{D_1}$ has the same form as $\mathbf{x}_{n,k,[m]}$. By the iterative processing, a posteriori probability of the received signals can be continuously maximized.

Table 1
Simulation parameters.

Carrier frequency: 2 GHz	Sampling frequency: 7.68 MHz
TTI duration: 1 ms	OFDM symbols per frame: 14
FFT size: 512	Subcarrier spacing: 15 kHz
System bandwidth: 5 MHz	RB: 12 subcarriers $\times$ 7 OFDM symbols
Code: Turbo	Received antenna: 2
Modulation	Desired cell: 16QAM
	Interference cell: 64QAM
Transmit data streams	Desired: 1
	Interference: one is strong and others are weak
Channel model [19,20]	Type: ETU
	Speed: 30 km/h
Proportional factor: ρ	3

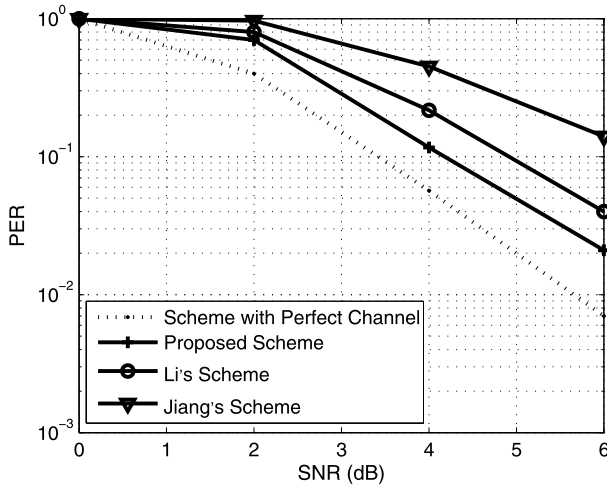


Fig. 5. The packet error ratio versus SNR when SIR = 4 dB.

4. Simulation results

In this section, we compare the performance of our scheme with that in [8–11]. For simplicity, we call the scheme in [8] and the scheme in [9–11] as Li's scheme and Jang's scheme, respectively. In our simulations, we exploit the three-cell system models and assume there is a strong one in the interferences from the adjacent cells while others are weak. Denote P_s and P_I as the power of transmit signal and the power of the strong interference, respectively, then we can define *signal-to-interference ratio* (SIR) as $SIR = \frac{P_s}{P_I}$. Meanwhile, 5 MHz system bandwidth, Turbo code, and QAM modulation are exploited. The detailed simulation parameters are listed in Table 1.

In Figs. 5 and 6, we show the *packet-error rate* (PER) versus the *signal-to-noise ratio* (SNR) when SIR is 4 dB and the PER versus SIR when SNR is 5 dB, respectively. From these figures, the proposed scheme has a better PER performance compared with Li's scheme and Jang's scheme. Meanwhile, the initial spatial covariance matrix in Li's scheme is estimated by the proposed scheme, this therefore makes Li's scheme to have a better performance than that of Jang's scheme. This also further verifies that the proposed scheme outperforms Jang's scheme. It should be noted that although the proposed algorithm may use the iterative process to continuously improve the estimation accuracy of the channel and the spatial covariance matrix, it will pay a price of increasing computation complexity compared with Li's scheme and Jang's scheme.

In Fig. 7, we show the PER performance versus the number of iterations when SIR is 4 dB. From this figure, the PER performance of the proposed scheme will decrease as the number of iterations increases.

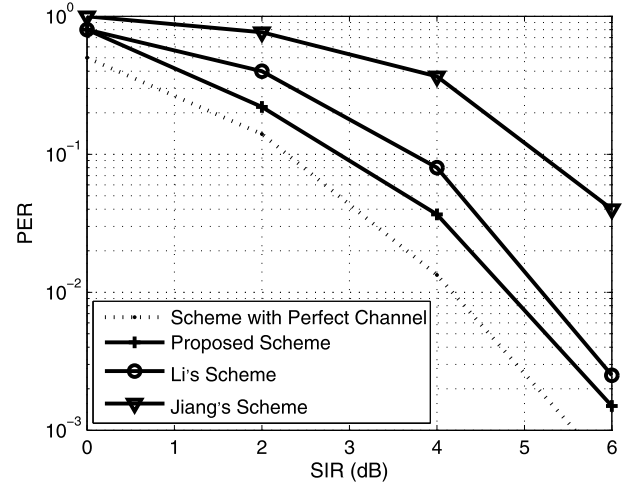


Fig. 6. The packet error ratio versus SIR when SNR = 5 dB.

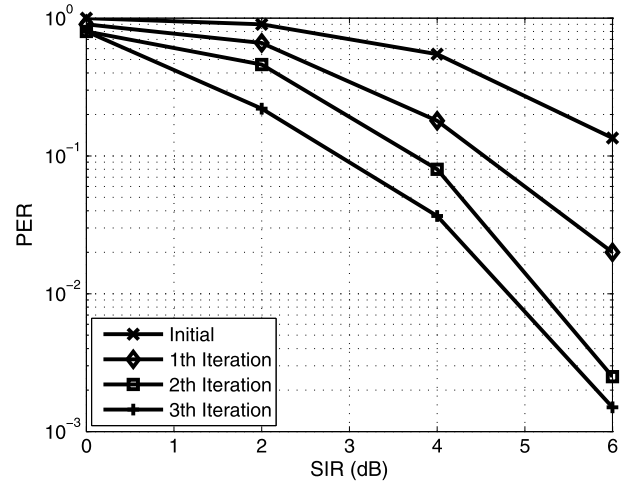


Fig. 7. The packet error ratio versus the number of iterations when SNR = 5 dB.

tions increases. Meanwhile, the PER performance of the proposed scheme with the high number of iterations will rapidly decrease as the SNR increases.

5. Conclusions

In this paper, we have proposed an iterative receiver with interference suppression for LTE downlink systems. It makes full use of the time–frequency correlation of channel to improve the estimation accuracy of channel and spatial covariance matrix. With the more accurate channel and spatial covariance matrix estimation, the capability of interference suppression and the successful probability of signal detection can be further improved by the maximum a posteriori detection. Since the proposed scheme is based on the pilot structure for LTE downlink systems and considers the practical constraints, it can be easily applied to receivers for LTE systems.

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References

- [1] S. Sesia, I. Toufik, M. Baker, *The UMTS Long Term Evolution: From Theory to Practice*, Wiley, 2009.
- [2] X. Wang, H.V. Poor, Iterative (turbo) soft interference cancellation and decoding for coded CDMA, *IEEE Trans. Commun.* 47 (7) (July 1999) 1046–1061.
- [3] R.C. de Lamare, R. Sampaio-Neto, Minimum mean-squared error iterative successive parallel arbitrated decision feedback detectors for DS-CDMA systems, *IEEE Trans. Commun.* 56 (5) (May 2008) 778–789.
- [4] H. Lee, B. Lee, I. Lee, Iterative detection and decoding with an improved V-BLAST for MIMO-OFDM Systems, *IEEE J. Sel. Areas Commun.* 24 (3) (March 2006) 504–513.
- [5] J.W. Choi, A.C. Singer, J. Lee, N.I. Cho, Improved linear soft-input soft-output detection via soft feedback successive interference cancellation, *IEEE Trans. Commun.* 58 (3) (March 2010) 986–996.
- [6] R.C. de Lamare, R. Sampaio-Neto, Adaptive reduced-rank equalization algorithms based on alternating optimization design techniques for MIMO systems, *IEEE Trans. Veh. Technol.* 60 (6) (July 2011) 2482–2494.
- [7] C. Kominakis, C. Fragouli, A.H. Sayed, R. Wesel, Multi-input multi-output fading channel tracking and equalization using Kalman estimation, *IEEE Trans. Signal Process.* 50 (5) (May 2002) 1065–1076.
- [8] Q. Li, J. Zhu, Q.H. Li, et al., Efficient spatial covariance estimation for asynchronous co-channel interference suppression in MIMO-OFDM systems, *IEEE Trans. Wirel. Commun.* 7 (12) (December 2008) 4849–4853.
- [9] J.H. Jang, J.S. Han, S.S. Kim, et al., Iterative receiver with enhanced spatial covariance matrix estimation in asynchronous interference environment for 3GPP LTE MIMO-OFDMA system, *IEEE Trans. Commun.* 92 (6) (2009).
- [10] J.H. Jang, J.S. Han, H.J. Choi, et al., Enhanced spatial covariance matrix estimation for asynchronous inter-cell interference mitigation in MIMO-OFDMA system, in: *IEEE 69th Vehicular Technology Conference, VTC Spring 2009*, 2009, pp. 1–5.
- [11] J.H. Jang, J.S. Han, S.S. Kim, et al., ML-based iterative channel and spatial covariance estimation in asynchronous interference environment for OFDMA system, in: *International Conference on Consumer Electronics, Digest of Technical Papers, ICCE'09*, 2009, pp. 1–2.
- [12] A. Jeremic, T.A. Thomas, A. Nehorai, OFDM channel estimation in the presence of interference, *IEEE Trans. Signal Process.* 52 (12) (December 2004) 3429–3439.
- [13] Y. Li, L.J. Cimini Jr., N.R. Sollenberger, Robust channel estimation for OFDM systems with rapid dispersive fading channels, *IEEE Trans. Commun.* 46 (7) (July 1998) 902–915.
- [14] Y. Li, N. Seshadri, S. Ariyavisitkul, Channel estimation for OFDM systems with transmitter diversity in mobile wireless channels, *IEEE J. Sel. Areas Commun.* 17 (3) (1999) 461–471.
- [15] 3GPP TS 36.211, Evolved universal terrestrial radio access (E-UTRA); physical channels and modulation (Release 8), 2009.
- [16] P. Hoeher, S. Kaiser, P. Robertson, Two-dimensional pilot-symbolaided channel estimation by Wiener filtering, in: *Proc. IEEE Intl. Conf. Acoustics, Speech, and Signal Processing, ICASSP'97*, 1997, pp. 1845–1848.
- [17] J. Ketonen, M. Juntti, J.R. Cavallaro, Performance-complexity comparison of receivers for a LTE MIMO-OFDM system, *IEEE Trans. Signal Process.* 58 (6) (June 2010) 3360–3372.
- [18] J. Ylioinas, M. Juntti, Iterative joint detection, decoding, and channel estimation in Turbo-coded MIMO-OFDM, *IEEE Trans. Veh. Technol.* 58 (4) (April 2009) 1784–1796.
- [19] Recommendation ITU-R M.1225, Guidelines for evaluation of radio transmission technologies for IMT-2000, Int. Telecommun. Union, 1997.
- [20] 3GPP TS 05.05 V8.15.0, Technical specification group GSM/EDGE radio access network; radio transmission and reception (release 1999), August 2001.

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